Study and Design of a Zero Voltage Switched Boost Converter

A THESIS SUBMITTED IN PARTIAL FULFILLMENT
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by

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CERTIFICATE

This is to certify that the thesis titled – "**Study and Design of a Zero Voltage Switched Boost Converter**" submitted by Sri **Rahul Shrivastava** and Sri **Gupta Saurabh** in partial fulfillment of the requirements for the award of Bachelor of Technology Degree in Electrical Engineering at the National Institute of Technology, Rourkela (Deemed University) is an authentic work carried out by them under my supervision and guidance. To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other University / Institute for the award of any Degree or Diploma.

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Date :

Rahul Shrivastava

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CONTENTS

ABSTRACT

We study theoretical circuit operation of zero voltage switching over the basic premise of boost converters (step-up dc chopper circuits). Zero-voltage switching technique is studied which, in contrast to zero-current switching, eliminates the switching loss and dv/dt noise due to the discharging of junction capacitances and the reverse recovery of diodes Zero Voltage Switching (ZVS) including various switching techniques in resonant converters is studied. Also a working model of a Zero Voltage Switched Boost Converter is constructed in the laboratory and its working and waveforms observed.

CHAPTER 1

INTRODUCTION

INTRODUCTION

Electronic power processing technology has evolved around two fundamentally different circuit schemes: duty-cycle modulation, commonly known as pulse width modulation (PWM), and resonance. The PWM technique processes power by interrupting the power flow and controlling the duty cycle, thus, resulting in pulsating current and voltage waveforms. The resonant technique processes power in a sinusoidal form. Due to circuit simplicity and ease of control, the PWM technique has been used predominantly in today's power electronics industries, particularly, in low-power power supply applications, and is quickly becoming a mature technology. Resonant technology, although well established in high-power SCR motor drives and uninterrupted power supplies, has not been widely used in low-power dc/dc converter applications due to its circuit complexity

With available devices and circuit technologies, PWM converters have been designed to operate generally at 30- 50-kHz switching frequency. In this frequency range, the equipment is deemed optimal in weight, size, efficiency, reliability and cost. In certain applications where high power density is of primary concern, the conversion frequency has been chosen as high as several hundred kilohertz. With the advent of power MOSFET'S, devices switching speed as high as tens of megahertz is possible. Accompanying the high switching frequency, however, are two major difficulties with the semiconductor devices, namely **high switching stress** and **switching loss.** For a given switching converter, the presence of leakage inductances in the transformer and junction capacitances in semiconductor devices causes the power devices to operate in inductive turn-off and capacitive turn-on. As the semiconductor device switches off an inductive load, voltage spikes are induced by the sharp di/dt across the leakage inductances, On the other hand, when the switch turns on at high voltage level, the energy stored in the device's output capacitances, 0.5 CV2, is trapped and dissipated inside the device. Furthermore, tum-on high voltage levels induces a severe switching noise known as the Miller effect which is coupled into the drive circuit, leading to significant noise and instability.

While not severe in lower switching frequencies, the capacitive tum-on loss due to the discharging of the parasitic junction capacitances of power MOSFET'S becomes the dominating factor when the switching frequency is raised to the megahertz range. For example, a junction capacitance of 100 pF, switching at 300 V, will induce a turn-on loss of 4.5 W at 1 MHz and 22.5 W at 5 MHz.

To overcome these drawbacks, the concept of Zero Current Switching Technique and Zero Voltage Switching have been introduced.

The paper presents the concept of Zero Current Switching Technique and Zero Voltage Switching Technique in detail. For the zero current switching technique, the objective is to use auxiliary ***LC*** resonant elements to shape the switching device's current waveform at on-time in order to create a zero-current condition for the device to turn off. The dual of the above statement is to use auxiliary ***LC*** resonant elements to shape the switching device's voltage waveform at off-time in order to create a zero-voltage condition for the device to turn on. This latter statement describes the principle of zero voltage switching. The recognition of the duality relationship between these two techniques leads to the development of the concept of voltage-mode resonant switches and a new family of converters operating under the zero-voltage switching principle.

CHAPTER 2

CHOPPER CIRCUITS

CHOPPER CIRCUITS

Many industrial applications require power from DC sources. Several of these applications, however, perform better in case these are fed from variable DC voltage sources. Examples of such DC system are subway cars, trolley buses, battery-operated vehicles, battery charging etc.

From an AC supply systems, variable DC output voltage can be obtained through the use of phase controlled converters or motor-generator sets. The conversion of fixed DC voltage to an adjustable DC output voltage. Through the use of semiconductor devices, can be carried out by the use of two types of DC to DC converters mentioned below.

(1) AC link chopper:

In the ac link chopper dc is first converted to ac by an inverter (dc to ac converter). AC is then stepped-up or stepped-down by a transformer which is then converted back to a dc by a diode rectifier. As the conversion is in two stages, dc to ac and then ac to dc, the link chopper is costly, bulky and less efficient

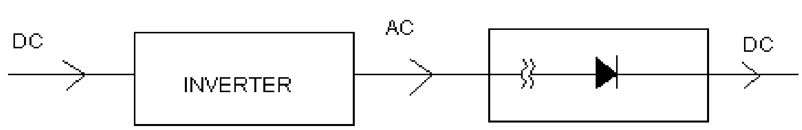


Fig 2.1 (a) AC link chopper

Fig 2.2(b) dc chopper

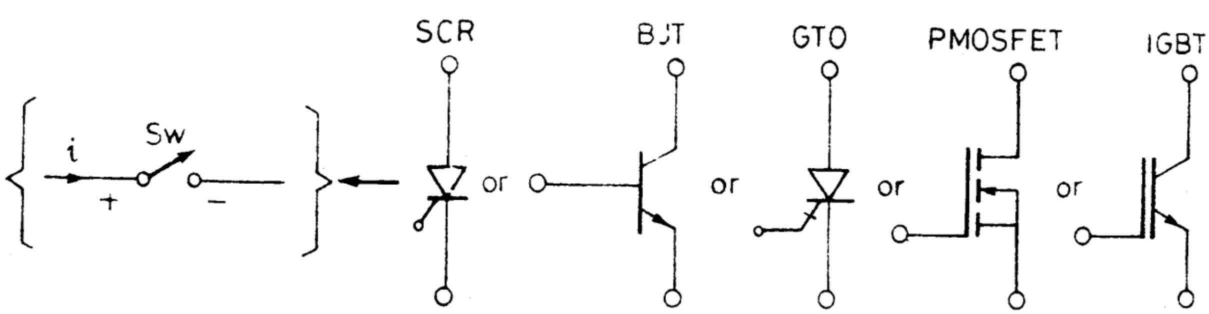


Fig 2.3 (c)Representation of power semiconductor device

(2) DC Chopper :

A chopper is a static device that converts fixed dc input voltage to a variable dc output voltage directly. A chopper may be thought of as dc equivalent of an ac transformer since they behave in an identical manner. As choppers involve one stage conversion, these are more efficient.

Choppers are now being used all over the world for rapid transit systems. These are also used in trolley cars, marine hoists etc. The future electric automobiles are likely to use choppers for their speed control and braking. Chopper systems offer smooth control, high efficiency, fast response and regeneration.

The power semiconductor devices used for a chopper circuit can be force-commutated thyristor, power BJT, power MOSFET, GTO or IGBT. Like the transformer, a chopper can also be used to step-down or step-up the fixed input voltage.

BOOST CONVERTER

A boost converter (step-up converter) is a power converter with an output DC voltage greater than its input DC voltage. It is a class of switching-mode power supply (SMPS) containing at least two semiconductor switches (a diode and a transistor) and at least one energy storage element. Filters made of capacitors (sometimes in combination with inductors) are normally added to the output of the converter to reduce output voltage ripple.

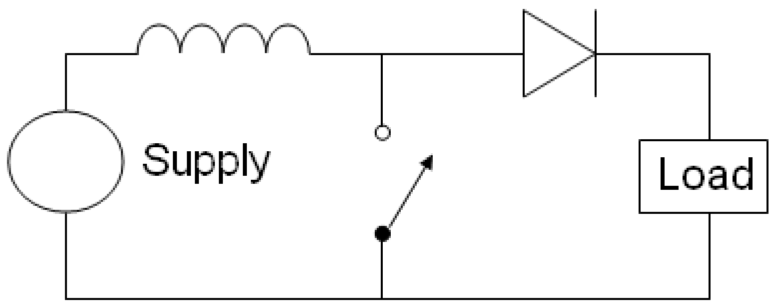


Fig 2.4 Basic schematic of a boost converter. The switch is typically a MOSFET, IGBT, or BJT.

Overview

Power can also come from DC sources such as batteries, solar panels, rectifiers, and DC generators. A process that changes one DC voltage to a different DC voltage is called DC to DC conversion. A boost converter is a DC to DC converter with an output voltage greater than the source voltage. A boost converter is sometimes called a step-up converter since it “steps up” the source voltage. Since power ($P = VI$) must be conserved, the output current is lower than the source current.

History

For high efficiency, the SMPS switch must turn on and off quickly and have low losses. The advent of a commercial semiconductor switch in the 1950’s represented a major milestone that made SMPSs such as the boost converter possible. Semiconductor switches turned on and off more quickly and lasted longer than other switches such as vacuum tubes and electromechanical relays. The major DC to DC converters were developed in the early-1960s when semiconductor switches had become available. The aerospace industry’s need for small, lightweight, and efficient power converters led to the converter’s rapid development.

Switched systems such as SMPS are a challenge to design since its model depends on whether a switch is opened or closed. R.D. Middlebrook from Caltech in 1977 published the models for DC to DC converters used today. Middlebrook averaged the circuit configurations for each switch state in a technique called state-space averaging. This simplification reduced two systems into one. The new model led to insightful design equations which helped SMPS growth.

Applications

Battery powered systems often stack cells in series to achieve higher voltage. However, sufficient stacking of cells is not possible in many high voltage applications due to lack of space. Boost converters can increase the voltage and reduce the number of cells. Two battery-powered applications that use boost converters are hybrid electric vehicles (HEV) and lighting systems.

The Toyota Prius HEV contains a motor which utilizes voltages of approximately 500 V. Without a boost converter, the Prius would need nearly 417 cells to power the motor. However, a Prius actually uses only 168 cells and boosts the battery voltage from 202 V to 500 V. Boost converters also power devices at smaller scale applications, such as portable lighting systems. A white LED typically requires 3.3V to emit light, and a boost converter can step up the voltage from a single 1.5 V alkaline cell to power the lamp. Boost converters can also produce higher voltages to operate cold cathode fluorescent tubes (CCFL) in devices such as LCD backlights and some flashlights.

CIRCUIT ANALYSIS

Operating principle

The key principle that drives the boost converter is the tendency of an inductor to resist changes in current. When being charged it acts as a load and absorbs energy (somewhat like a resistor), when being discharged, it acts as an energy source (somewhat like a battery). The voltage it produces during the discharge phase is related to the rate of change of current, and not to the original charging voltage, thus allowing different input and output voltages.

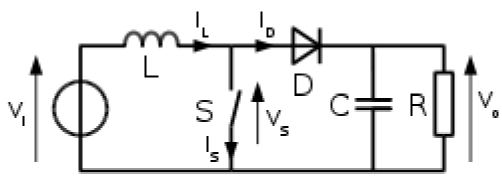


Fig 2.5 Boost converter schematic

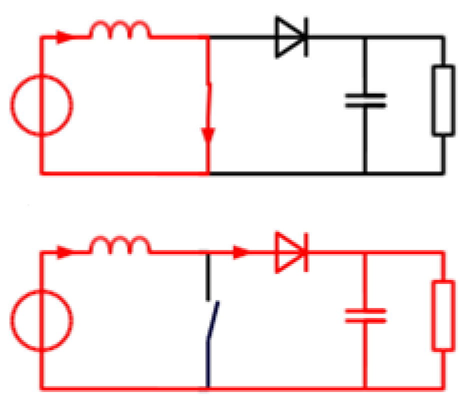


Fig 2.6 The two configurations of a boost converter, depending on the state of the switch S.

The basic principle of a Boost converter consists in 2.6 distinct states (figure 2.6)

- in the On-state, the switch S (see figure 2.6) is closed, resulting in an increase in the inductor current;
- in the Off-state, the switch is open and the only path offered to inductor current is through the flyback diode D, the capacitor C and the load R. This results in transferring the energy accumulated during the On-state into the capacitor.

- The input current is the same as the inductor current as can be seen in figure 2.6. So it is not discontinuous as in the buck converter and the requirements on the input filter are relaxed compared to a buck converter.

Continuous mode

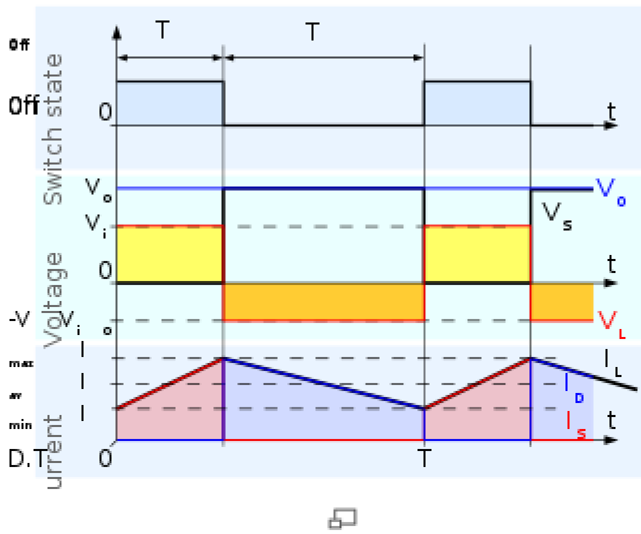


Fig 2.7 Waveforms of current and voltage in a boost converter operating in continuous mode.

When a boost converter operates in continuous mode, the current through the inductor (I_L) never falls to zero. Figure 2.7 shows the typical waveforms of currents and voltages in a converter operating in this mode. The output voltage can be calculated as follows, in the case of an ideal converter (i.e using components with an ideal behaviour) operating in steady conditions:

During the On-state, the switch S is closed, which makes the input voltage (V_i) appear across the inductor, which causes a change in current (I_L) flowing through the inductor during a time period (t) by the formula:

$$\frac{\Delta I_L}{\Delta t} = \frac{V_i}{L}$$

At the end of the On-state, the increase of I_L is therefore:

$$\Delta I_{L_{On}} = \frac{1}{L}\int_0^{DT} V_i dt = \frac{DT}{L} V_i$$

D is the duty cycle. It represents the fraction of the commutation period T during which the switch is On. Therefore D ranges between 0 (S is never on) and 1 (S is always on).

During the Off-state, the switch S is open, so the inductor current flows through the load. If we consider zero voltage drop in the diode, and a capacitor large enough for its voltage to remain constant, the evolution of I_L is:

$$V_i - V_o = L\frac{dI_L}{dt}$$

Therefore, the variation of I_L during the Off-period is:

$$\Delta I_{L_{Off}} = \int_0^{(1-D)T} \frac{(V_i - V_o)\,dt}{L} = \frac{(V_i - V_o)\,(1-D)\,T}{L}$$

As we consider that the converter operates in steady-state conditions, the amount of energy stored in each of its components has to be the same at the beginning and at the end of a commutation cycle. In particular, the energy stored in the inductor is given by:

$$E = \frac{1}{2}LI_L^2$$

Therefore, the inductor current has to be the same at the beginning and the end of the commutation cycle. This can be written as

$$\Delta I_{L_{On}} + \Delta I_{L_{Off}} = 0$$

Substituting $\Delta I_{L_{On}}$and $\Delta I_{L_{Off}}$by their expressions yields:

$$\Delta I_{L_{On}} + \Delta I_{L_{Off}} = \frac{V_i DT}{L} + \frac{(V_i - V_o)\,(1-D)\,T}{L} = 0$$

This can be written as:

$$\frac{V_o}{V_i} = \frac{1}{1-D}$$

Which in turns reveals the duty cycle to be:

$$D = 1 - \frac{V_i}{V_o}$$

From the above expression it can be seen that the output voltage is always higher than the input voltage (as the duty cycle goes from 0 to 1), and that it increases with D, theoretically to infinity as D approaches 1. This is why this converter is sometimes referred to as a *step-up* converter.

Discontinuous mode

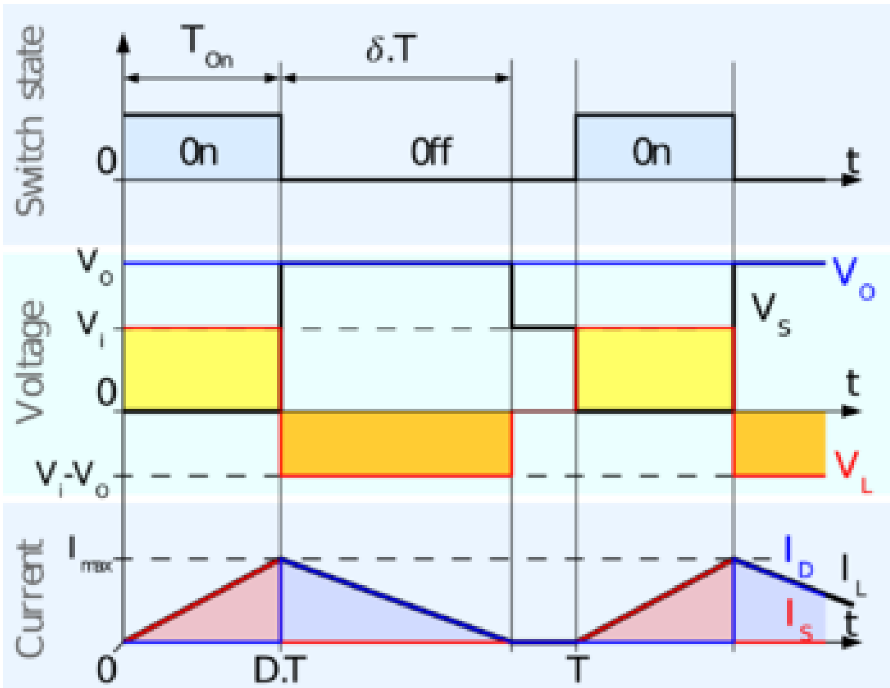


Fig 2.8 Waveforms of current and voltage in a boost converter operating in discontinuous mode.

In some cases, the amount of energy required by the load is small enough to be transferred in a time smaller than the whole commutation period. In this case, the current through the inductor falls to zero during part of the period. The only difference in the principle described above is that the inductor is completely discharged at the end of the commutation cycle (see waveforms in figure 2.8). Although slight, the difference has a strong effect on the output voltage equation. It can be calculated as follows:

As the inductor current at the beginning of the cycle is zero, its maximum value $I_{L_{Max}}$(at *t* = *DT*) is

$$I_{L_{Max}} = \frac{V_i DT}{L}$$

During the off-period, I_L falls to zero after δT:

$$I_{L_{Max}} + \frac{(V_i - V_o)\,\delta T}{L} = 0$$

Using the two previous equations, δ is:

$$\delta = \frac{V_i D}{V_o - V_i}$$

The load current I_o is equal to the average diode current (I_D). As can be seen on figure 4, the diode current is equal to the inductor current during the off-state. Therefore the output current can be written as:

$$I_o = \bar{I_D} = \frac{I_{L_{max}}}{2}\delta$$

Replacing I_{Lmax} and δ by their respective expressions yields:

$$I_o = \frac{V_i D T}{2L} \cdot \frac{V_i D}{V_o - V_i} = \frac{V_i^2 D^2 T}{2L\,(V_o - V_i)}$$

Therefore, the output voltage gain can be written as flow:

$$\frac{V_o}{V_i} = 1 + \frac{V_i D^2 T}{2 L I_o}$$

Compared to the expression of the output voltage for the continuous mode, this expression is much more complicated. Furthermore, in discontinuous operation, the output voltage gain not only depends on the duty cycle, but also on the inductor value, the input voltage, the switching frequency, and the output current.

CHAPTER 3

CONTROL STRATEGIES

CONTROL STRATEGIES:

It is seen that the average value of output voltage V_0 can be controlled through D by opening and closing the semiconductor switch periodically. The various control strategies for varying the duty cycle D are as follows:

1. Time ratio control (TRC)
2. Current-limit control

1. Time Ratio Control (TRC):

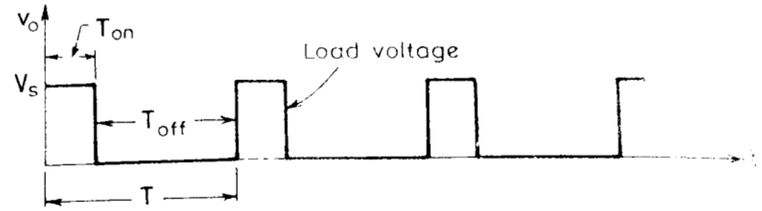
As the name suggests in this control scheme the duty cycle is varied. This is realized in two different control strategies :

(a) Constant frequency system
(b) Variable frequence system

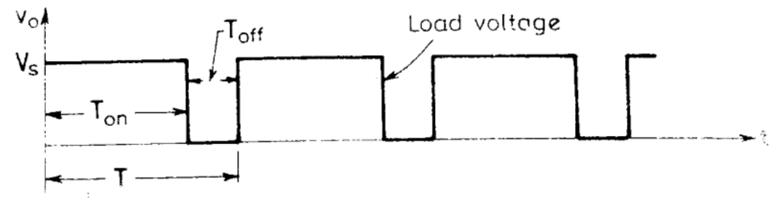
(a) **Constant frequency system:** In this scheme the on time T_{on} is varied but the chopping frequency f (or the chopping period T) is kept constant. Variation of T_{on} means adjustment of pulse width; as such this scheme is also called **Pulse Width Modulation (PWM)** Scheme.

(b) **Variable frequency system**: In this scheme the chopping frequency f (or chopping period T) is varied and either on time T_{on} or off time T_{off} is kept constant. This method of controlling D is also called **Frequency Modulation Scheme.**

It is seen that PWM scheme is better than the variable frequency scheme. PWM technique however has a limitation as T_{on} can not be reduced to near-zero for most of the commutation circuits used in choppers. As such low range of D (duty cycle) control is not possible in PWM. However this can be achieved by increasing the chopping period (decreasing the chopper frequency) of the chopper.

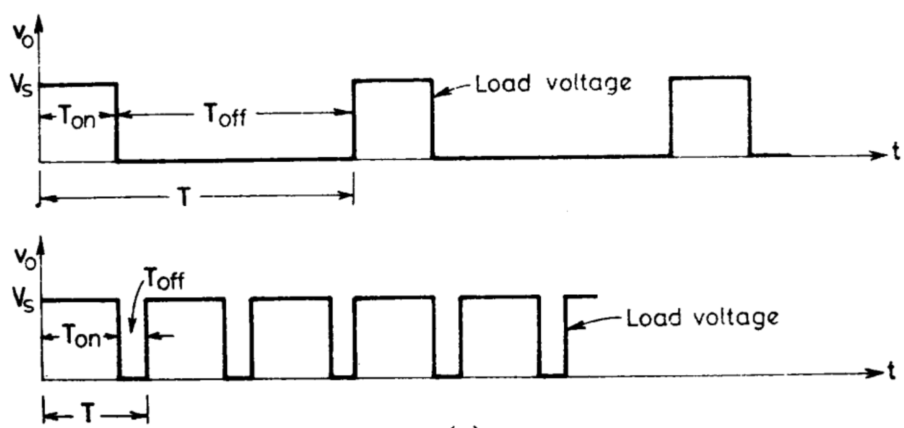


(a)

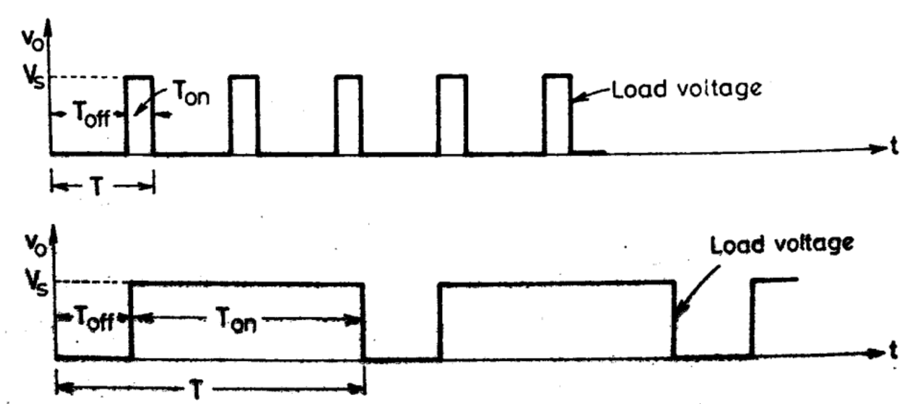


(b)

Principle of pulse-width modulation (constant T).



(a)



(b)

Fig 3.1 (a) on-time Ton constant (b) off-time Toff constant

2. Current Limit Control:

In this strategy, the on and off chopper circuit is guided by the previous set values of load current. These two set values are maximum load current $I_{0.max}$ and minimum load current $I_{0.min}$.

When load current reaches maximum limit the chopper is switched off. Now load current free-wheels and begins to decay exponentially. When it falls to lower limit (minimum value), chopper is switched on and load current begins to rise as shown.

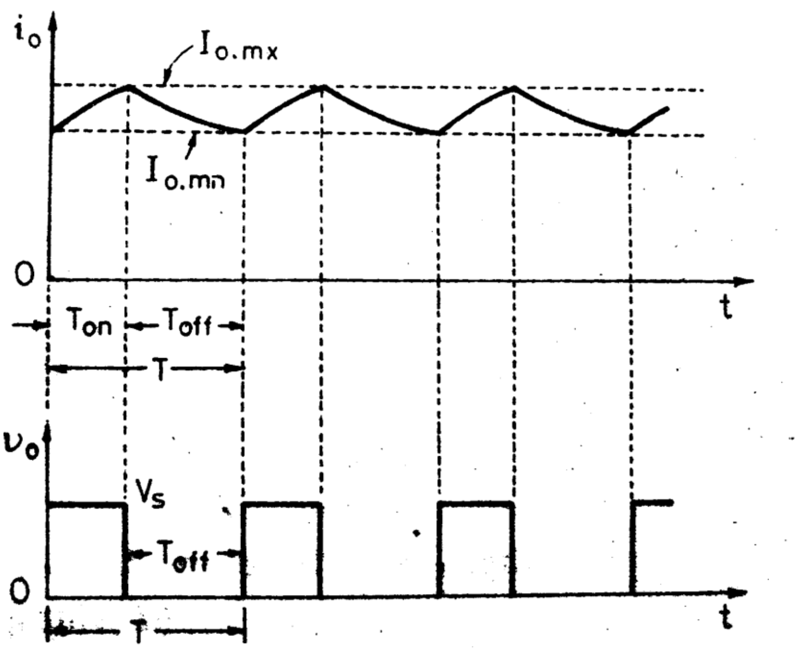


Fig 3.2 Current Limit Control for Chopper

Switching frequency of chopper can be controlled by using $I_{0.max}$ and $I_{0.min.}$ Ripple current (= $I_{0.max}$ - $I_{0.min}$) can be lowered and this in turn necessitates higher switching frequency and therefore more switching losses.

Circuit limit control involves a feedback loop the trigger circuitry for chopper is therefore more complex.

PWM technique is, therefore, the commonly chosen control strategy of the power control chopper circuit.

CHAPTER 4

ZERO CURRENT SWITCHING RESONANT CONVERTERS

ZERO CURRENT SWITCHING RESONANT CONVERTERS

The switches of Zero Current Switching (ZCS) resonant converters turn on and off at zero current. The resonant circuit that consists f switch S_1, inductor L, and capacitor C is shown. The inductor L is connected in series with power switch S_1 to achieve ZCS. It is classified into two types – L type and M type. In both the types the inductor L limits the di/dt of the switch current and L and C constitute a series resonant circuit. When the switch current is zero there is a current $i=C_j.dv_t/dt$ flowing through the internal capacitance C_j due to finite slope of switch voltage at turn off. This current flow causes power dissipation in the switch and limits the high switching frequency.

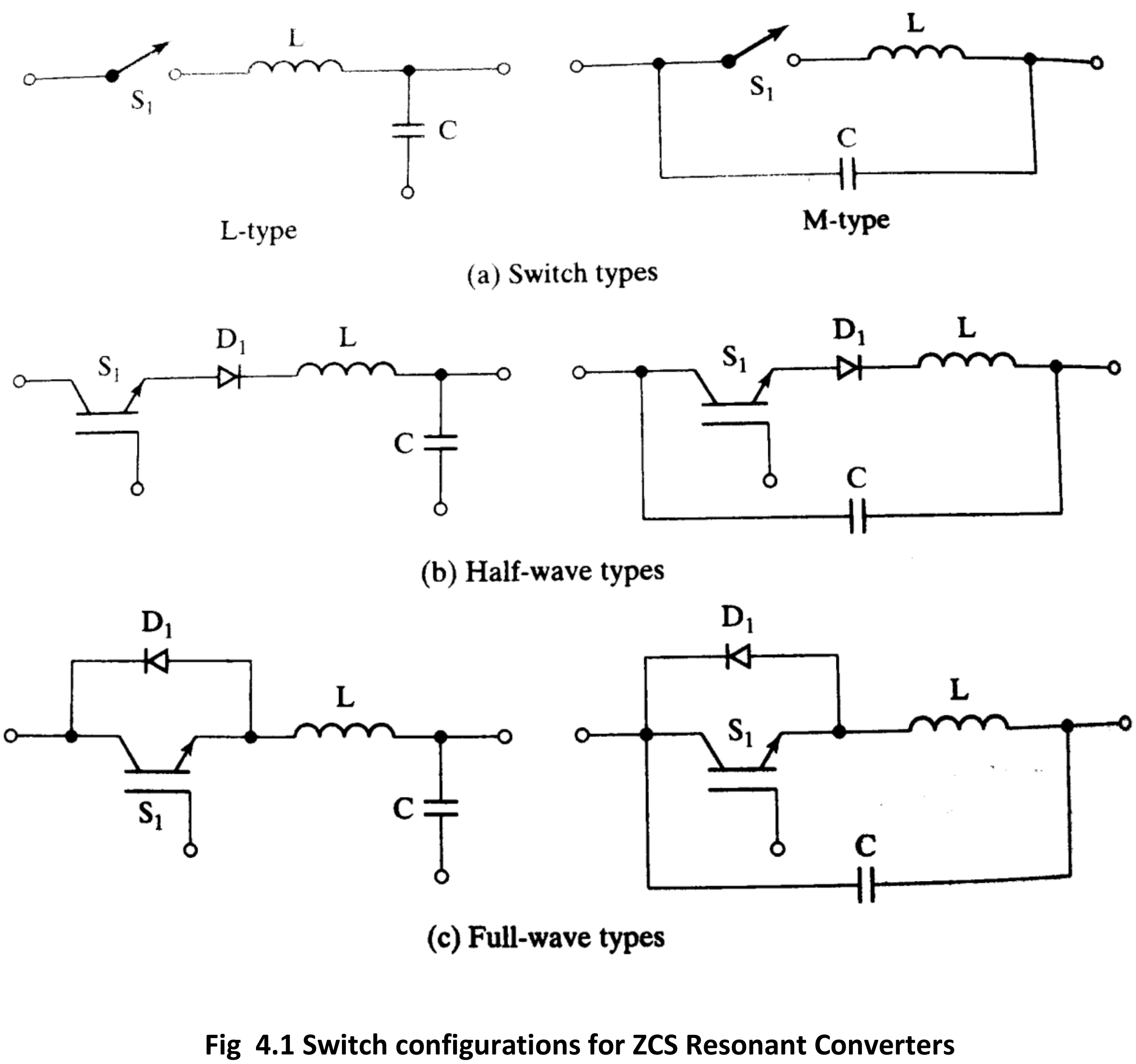


Fig 4.1 Switch configurations for ZCS Resonant Converters

The switch can be implemented either in half wave configuration where diode D_1 allows unidirectional current flow or in full-wave configuration where the switch current can flow bidirectionally. The practical devices do not turn off at zero current due to their recovery time. As a result, an amount of energy can be trapped in inductor L of the m-type configuration and voltage transients appear across the switch. This normally favors L type configuration over M type one.

L-TYPE ZCS RESONANT CONVERTER

The circuit operation can be divided into 5 modes whose equivalent circuits are shown. We shall redefine the time origin, t=0, at the beginning of each mode.

Mode 1 – This mode is valid for $0 <= t <= t_{1.}$ Switch S_1 is turned on and diode D_m conducts. The inductor current i_L which rises linearly is given by

$$I_L = (V_s/L)t$$

This mode ends at time $t=t_1$ when i_l $(t=t_1) = I_0$. That is $t1 = I_0L/V_s$,

Mode 2 – This mode is valid for $0 <= t <= t_2$. Switch S_1 remains on but diode Dm is off. The inductor current iL is given by

$$i_L=I_m\sin \omega_o t + I_0$$

where $I_m = V_s\surd(C/L)$ and $\omega_0 = 1/\surd LC$. The capacitor voltage v_c is given by

$$v_c = V_s(1 – \cos\omega_0 t)$$

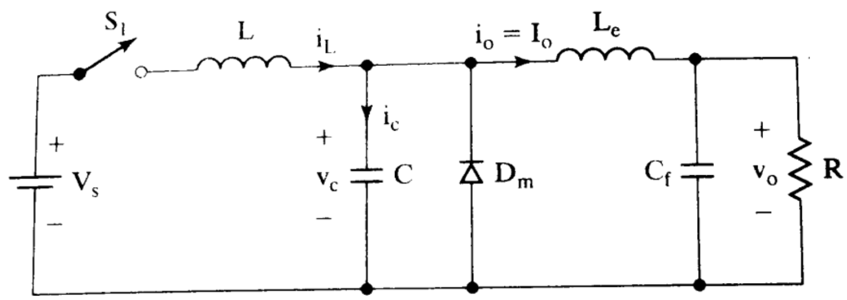
The peak current which occurs at $t = (\pi/2)\surd LC$ is

$$I_p = I_m + I_0$$

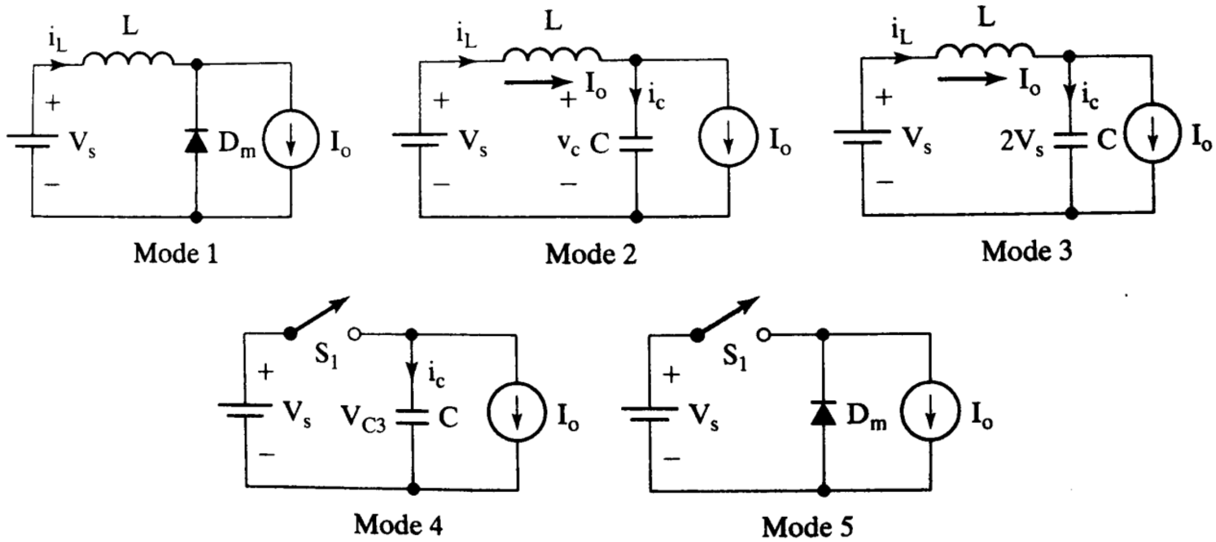
The peak capacitor voltage is given by

$$V_{c(pk)} = 2V_s$$

This mode ends at $t = t_2$ when $i_L(t = t_2) = I_0$ and $v_c(t = t_2) = V_{c2} = 2V_s$. Therefore $t_2 = \pi\surd LC$.



(a) Circuit



(b) Equivalent circuits

Mode 3 – This mode is valid for 0 <= t <=t_3. The inductor current that falls from I0 to zero is given by

$$i_L = I_0 - Im\sin\omega_0 t$$

The capacitor voltage is given by

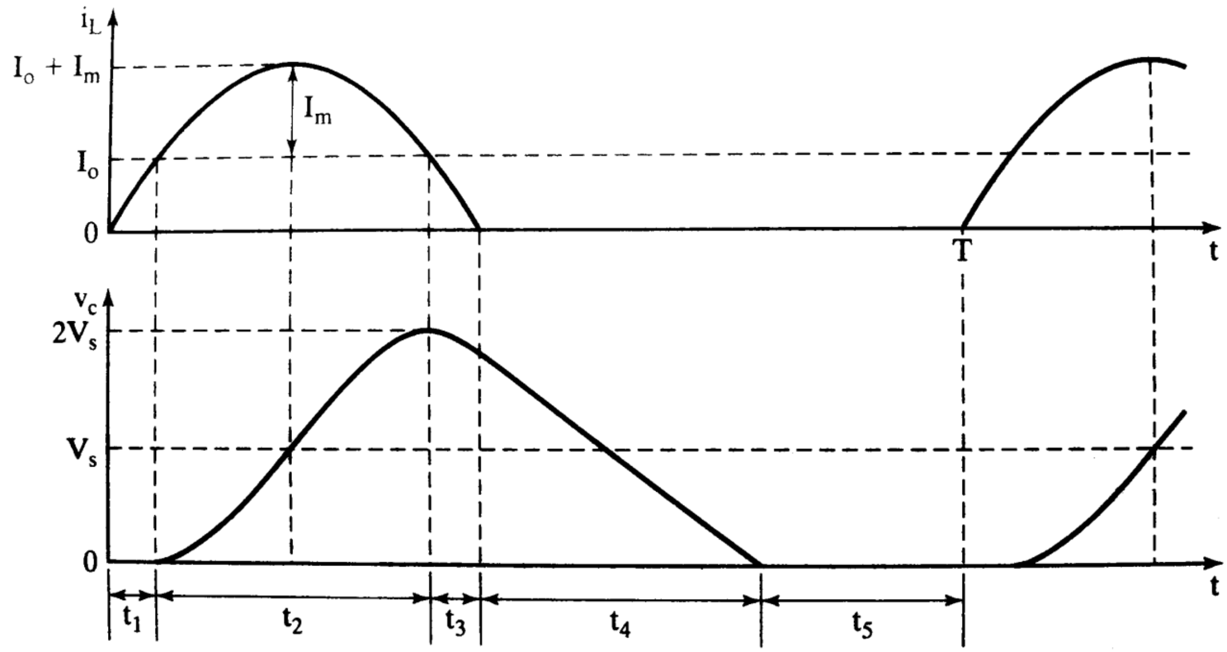
$$V_c = 2Vs\cos\omega_0 t$$

This mode ends at t = t_3 when i_L (t = t_3) = 0. And v_c(t = t_3) = V_{c3}. Thus $t_3 = \sqrt{LC}\sin^{-1}(1/x)$ where x = I_m/I_0 = = $(v_s/I_0)\sqrt{C/L}$

Mode 4 – This mode is valid for 0 <= t <= t_4. The capacitor supplies the load current Io and its voltage is given by

$$V_c = V_{c3} - (I_0/C)t$$

This mode ends at t = t_4 when vc (t = t_4) = 0. Thus $t_4 = V_{c3}C/I_0$



(c) Waveforms

L-Type ZCS Resonant Converter

Mode 5 – This mode is valid for $0 <= t <= t_5$. When the capacitor voltage tends to be negative, the diode D_m conducts. The load current I_o flows through diode Dm. This mode ends at time $t=t_5$ when the switch S_1 is turned on again and the cycle is repeated i.e. $t_5=T-(t_1+t_2+t_3+t_4)$

The wavefor for I_l and V_c are shown. The peak switch voltage equals the dc supply voltage. Because the switch current is zero at turn on and turn off, the switching loss, which is the product of v and I, becomes very small. The peak resonant current I_m must be higher than the load current I_o and this sets a limit on the minimum value of load resistance R. However by placing an anti-parallel diode across the switch the output voltage can be made insensitive to load variations.

CHAPTER 5

ZERO VOLTAGE SWITCHING RESONANT CONVERTERS

ZERO-VOLTAGE-SWITCHING RESONANT CONVERTERS

The switches of ZVS resonant converters turn on and off at zero voltage.

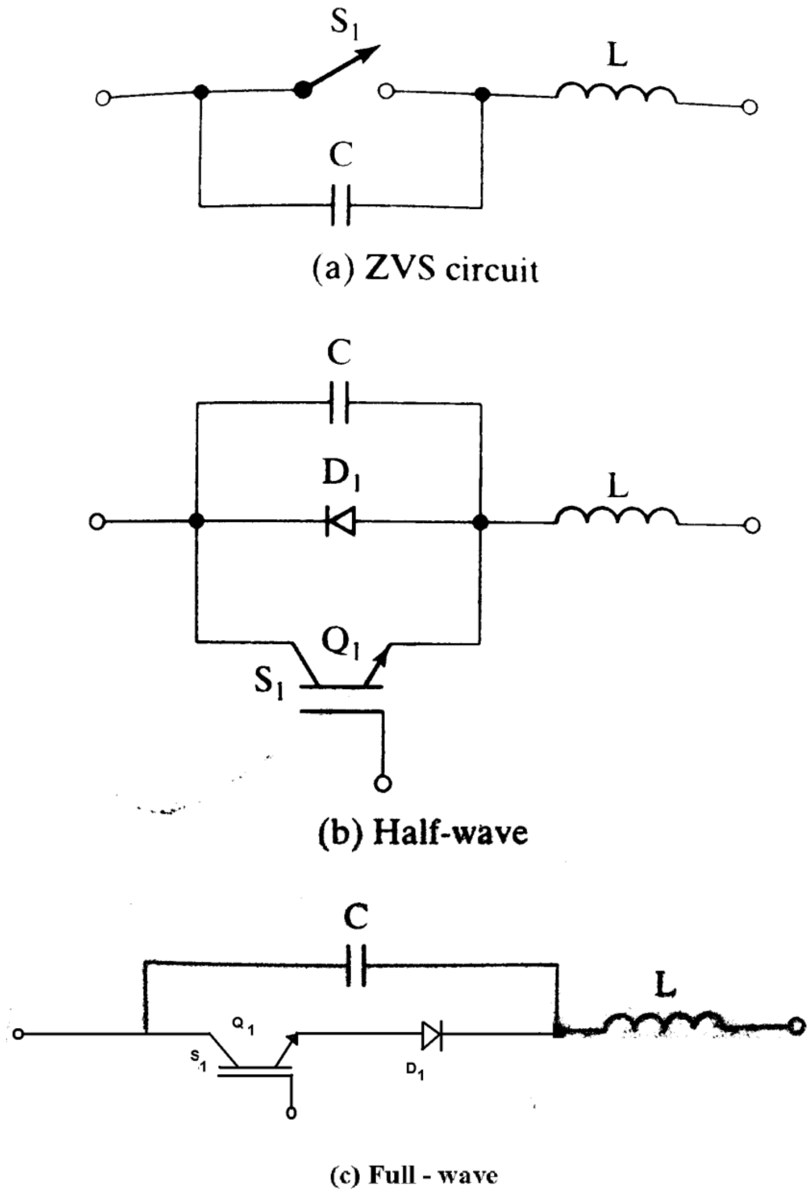


Fig 5.1 Switch Configurations for ZVS Resonant Converters

The capacitor C is connected in parallel with the switch S_1 to achieve ZVS. The internal switch capacitance C_j is added with the capacitor C and it affects the resonant frequency only, thereby contributing no power dissipation in the switch. If the switch is implemented with transistor Q_1 and an anti-parallel diode D_1 as shown, the voltage across C is clamped by D_1 and the switch is operated in half wave configuration. If the diode D_1 is connected in series with Q_1 as shown, the voltage across C can oscillate freely and the switch is operated in full wave configuration. A ZVS resonant converter is shown. A ZVS resonant converter is the dual of ZCS resonant converter.

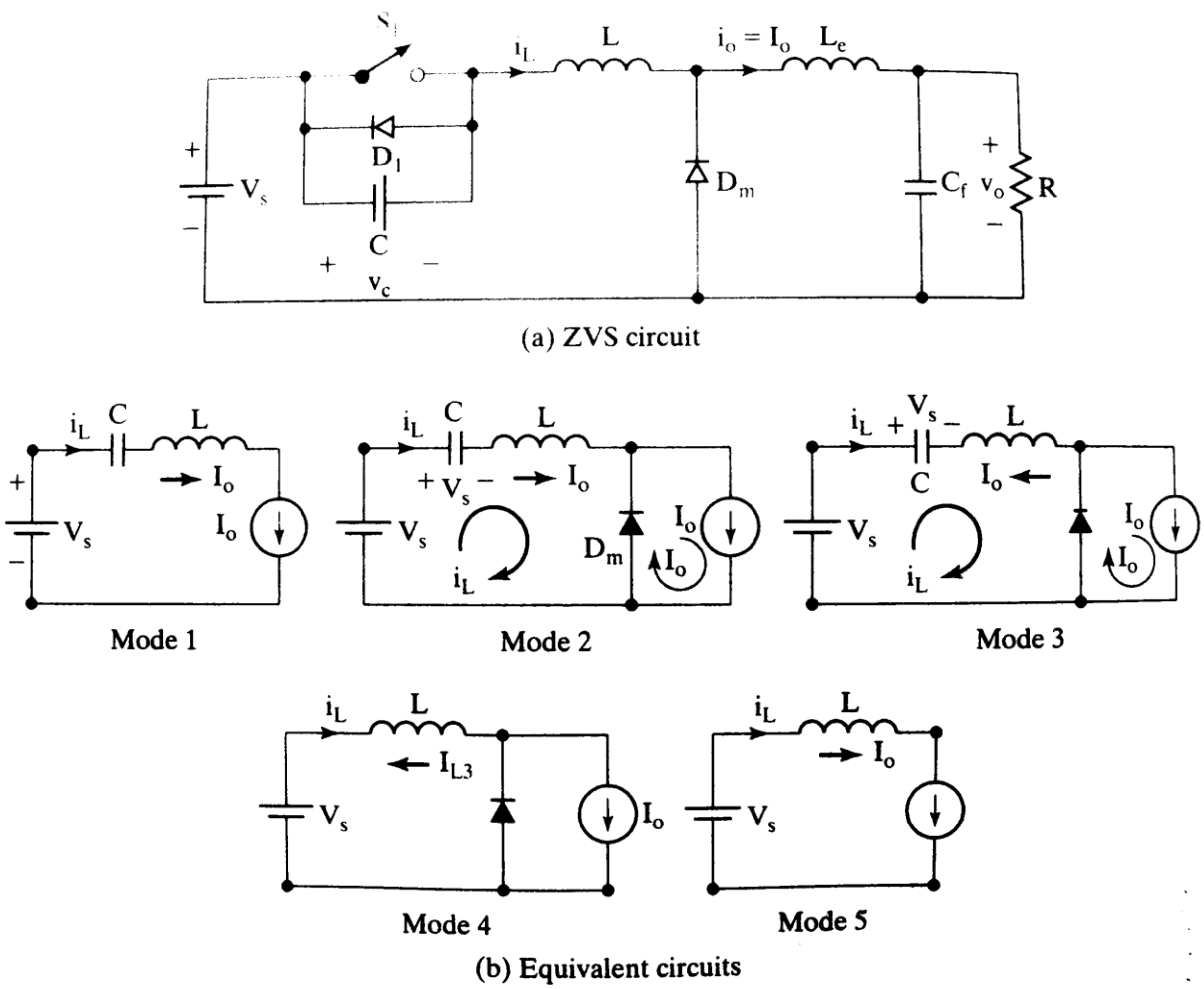
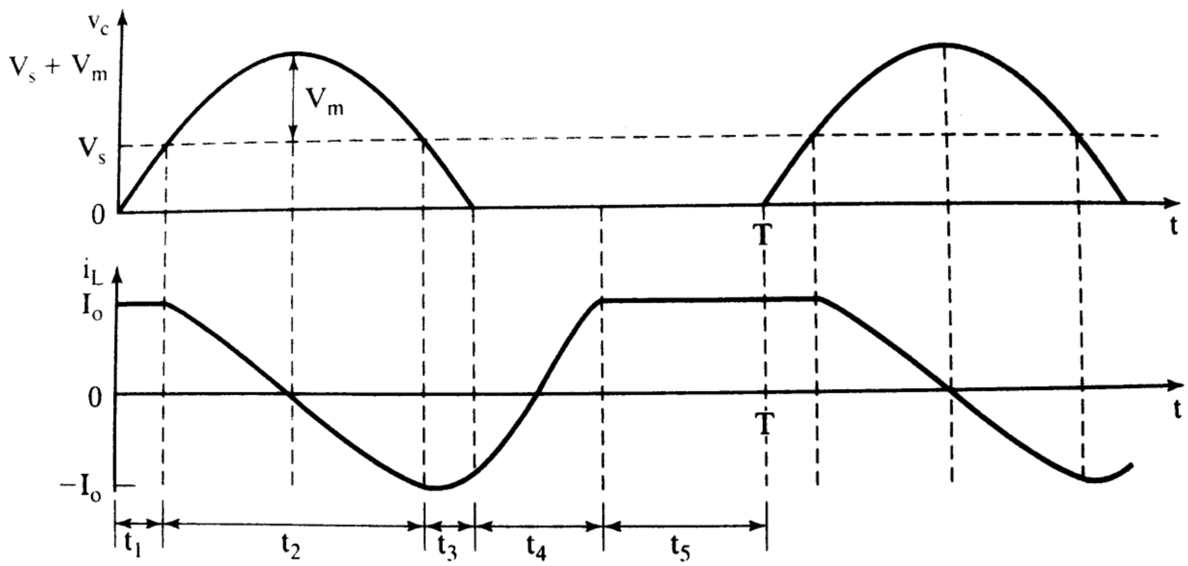


Fig 5.2



(c) Waveforms

Fig 5.3 ZVS Resonant Converter

The circuit operation can be divide in to 5 modes whose circuits are shown. We shall redefine the time origin, t=0, at the beginning of each mode.

Mode 1 : This mode is valid for $0 \le t \le t_1$. Both switch S_1 and diode D_m are off. Capacitor C charges at a constant rate of load current I_0. The capacitor voltage v_c which rises is given by

$$V_c = I_{o.t} / C$$

This mode ends at time $t = t_1$ when v_c ($t = t_1$) = V_s. That is $t_1 = V_s.C / I_0$.

Mode 2 : This mode is valid for $0 \le t \le t_2$. The switch S_1 is still off, but diode D_m turns on. The capacitor voltate v_c is given by

$$V_c = V_m \sin \omega t + V_s$$

Where $V_m = I_0\sqrt{(L/C)}$. The peak switch voltage which occurs at $t = (\pi/2) \sqrt{LC}$, is

$$V_{t(pk)} = V_{c(pk)} = I_0 \sqrt{(L/C)} + V_s$$

The inductor current i_L is given by

$$i_L = I_0 \cos \omega_0 t$$

This mode ends at $t = t_2$ when v_c $(t = t_2) = V_s$, and $i_L(t = t_2) = -I_0$. Therefore, $t_2 = \pi\sqrt{LC}$.

Mode 3 : This mode is valid for $0 \le t \le t_3$. The capacitor voltage that falls from v_s to zero is given by

$$V_c = V_s - V_m \sin \omega_0 t$$

The inductor current i_L is given by

$$i_L = -I_0 \cos \omega_0 t$$

This mode ends at $t = t_3$ when v_c $(t = t_3) = 0$, and i_L $(t = t3) = i_{L3}$. Thus,

$$T_3 = \sqrt{(LC)} \sin^{-1} x$$

Where, $x = V_s/V_m = (V_s/I_0)\sqrt{(C/L)}$.

Mode 4 : This mode is valid for $0 \le t \le t_4$. Switch S_1 is turned on and diode D_m remains on. The inductor current which rises linearly from I_{l3} to I_0 is given by

$$i_L = I_{L3} + (V_s/L)t$$

This mode ends at time $t = t_4$ when i_L $(t = t_4) = 0$. Thus $t_4 = (I_0 - I_{L3})(L/V_s)$. I_{L3} has a negative value.

Mode 5 : This mode is valid for $0 \le t \le t_5$. Switch S_1 is on but D_m is off. The load current I_0 flows through the switch. This mode ends at time $t = t_5$, when the switch S_1 is turned off again and the cycle is repeated. That is $t_5 = T - (t_1 + t_2 + t_3 + t_4)$.

The waveforms for i_L and v_c are shown. The equation

$$V_{t(pk)} = V_{c(pk)} = I_0 \sqrt{(L/C)} + V_s$$

shows that the peak switch voltage $V_{t(pk)}$ is dependent on the load current I_0. Therefore a wide variation in the load current results in a wide variation of the switch voltage. For this reason, ZVS converters are used only for constant-load applications. The switch must be turned on only at zero voltage. Otherwise, the energy stored in C can be dissipated in the switch. To avoid this situation, the antiparallel diode D_1 must conduct before turning on the switch.

CHAPTER 6

COMPARISON OF ZCS AND ZVS RESONANT CONVERTERS

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SWITCHING TECHNIQUES

COMPARISON OF ZCS AND ZVS RESONANT CONVERTERS

ZCS can eliminate switching losses at turnoff and reduce the switching losses at turnon. Because a relatively large capacitor is connected across the diode D_m the inverter operation becomes insensitive to the diode's junction capacitance. When power MOSFETs are used for ZCS the energy stored in the device's capacitance is dissipated during turn on. This capacitive turn on loss is proportional to the switching frequency.

During turn on a high rate of change of voltage may appear in the gate drive circuit due to the coupling through the Miller capacitor, thus increasing switching loss and noise. Another limitation is that the switches are under high current stress, resulting in higher conduction loss.

By the nature of ZCS, the peak switch current is much higher. In addition, a high voltage becomes established across the switch in the off state after the resonant oscillation. When the switch is turned on again, the energy stored in the output capacitor becomes discharged through the switch, causing a significant power loss at high frequency and higher voltages. This switching loss can be reduced by using ZVS.

ZVS eliminates the capacitive turn on loss. It is suited for high frequency operation. Without any voltage clamping, the switches may be subjected to excessive voltage stress which is proportional to the load and the output voltage can be achieved by varying the frequency.

ZERO-VOLTAGE SWITCHING TECHNIQUE

Voltage-Mode Resonant Switches A resonant switch represents a sub-circuit consisting of semiconductor switch SI and auxiliary resonant elements L, and C,. For a current-mode resonant switch, as shown in Fig. l(a), inductor L, is in series with switch SI to achieve zero-current switching; in a voltage-mode resonant switch, as shown in Fig. l(b), capacitor C, is in parallel with switch SI to achieve zero-voltage switching. As in the case of a current-mode resonant switch, the structure of SI determines the operation mode of the voltage- mode resonant switch. If the ideal switch SI is implemented by a transistor Q, and an anti-parallel diode, D,, as shown in Fig. 2(b), the voltage across capacitor C, is clamped by DI to a minimum value, and the resonant switch is operating in a half-wave mode. On the other hand, if SI is implemented by Q, in series with D,, as shown in Fig. 2(c), and the voltage on C, can oscillate freely, then the resonant switch is operating in a full-wave mode. Notice that in a current-mode resonant switch, the resonant interaction between L, and C, is initiated by the turn-on of SI, while on a voltage-mode resonant switch, it is initiated by the turn-off of SI.

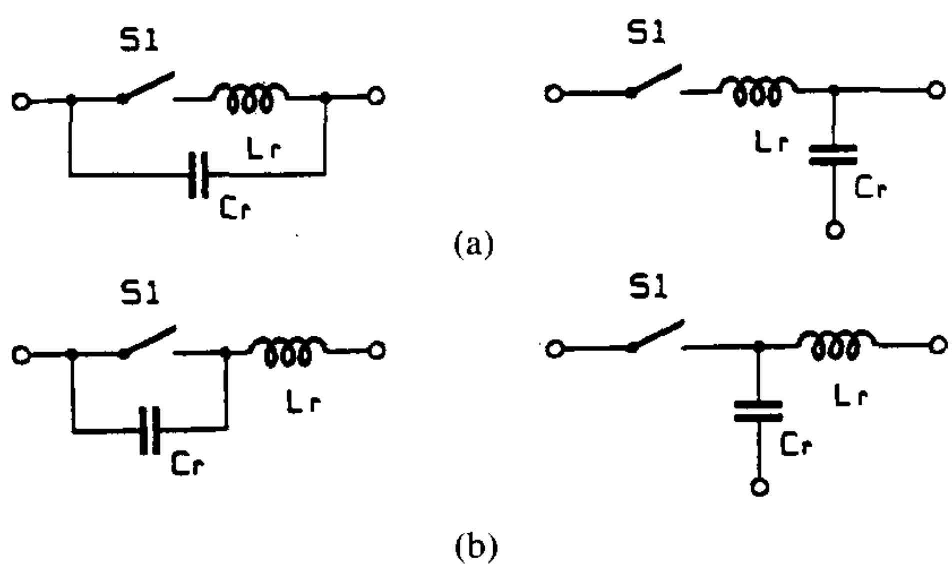

Fig 6.1. (a) Current-mode resonant switches. (b) Voltage-mode resonant switches

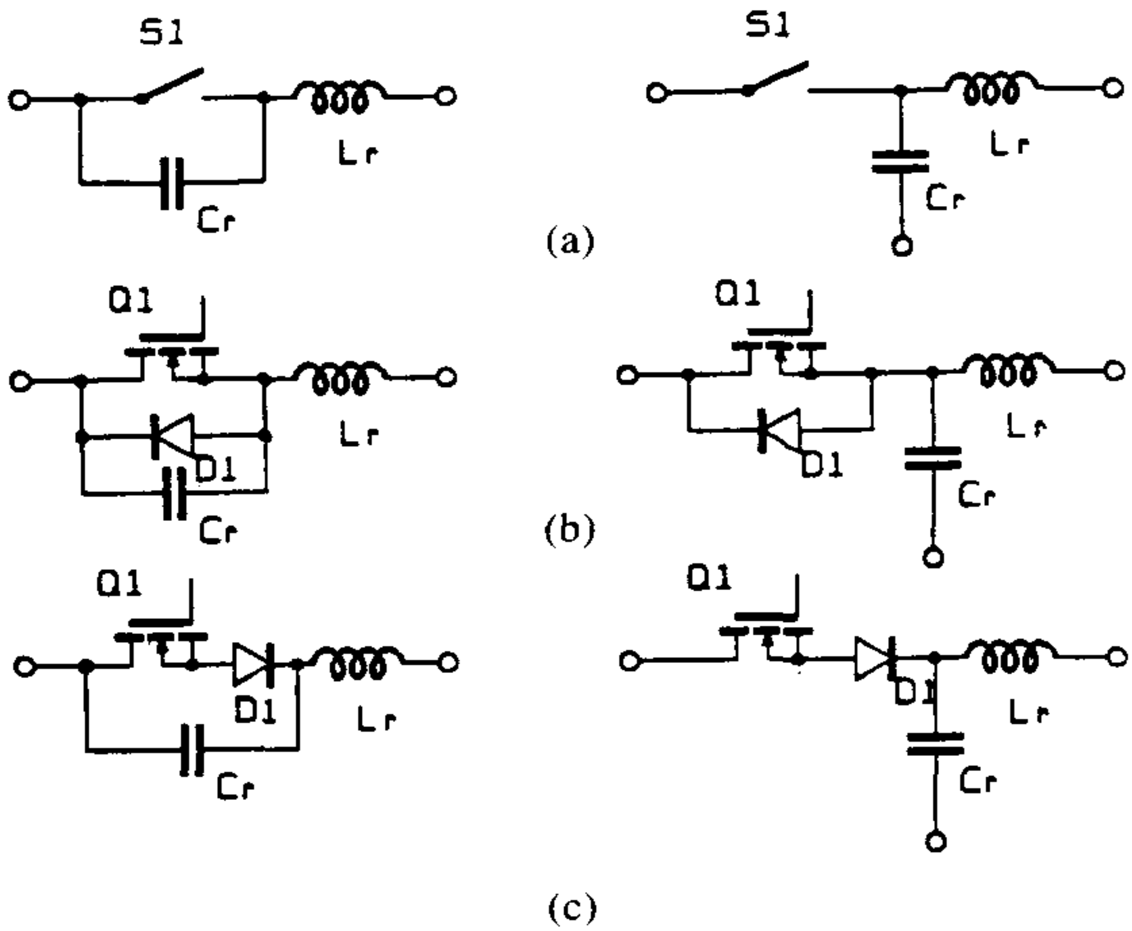

Fig 6.2. Voltage-mode resonant switches. (a) General notation. (b) Halfwave mode implementation. (c) Full-wave mode implementation.

CHAPTER 7

CONSTRUCTION PROJECT

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OBSERVATIONS

CIRCUIT DIAGRAM OF THE ZVS BOOST CONVERTER

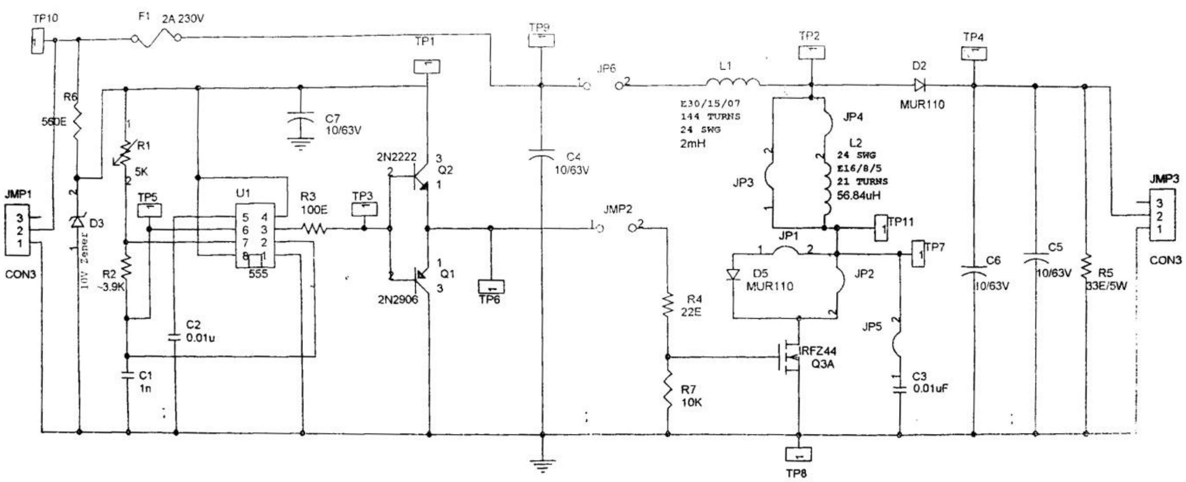


Fig 7.1 7.5W ZVS Boost Converter Circuit

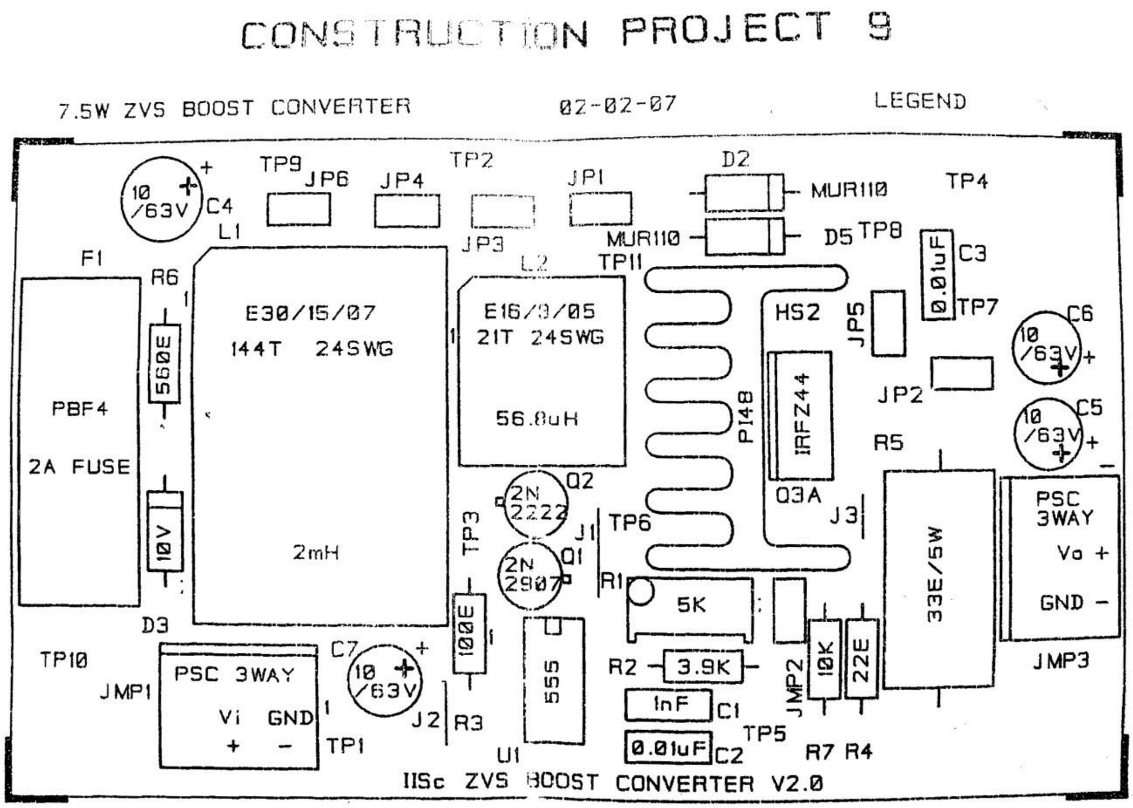


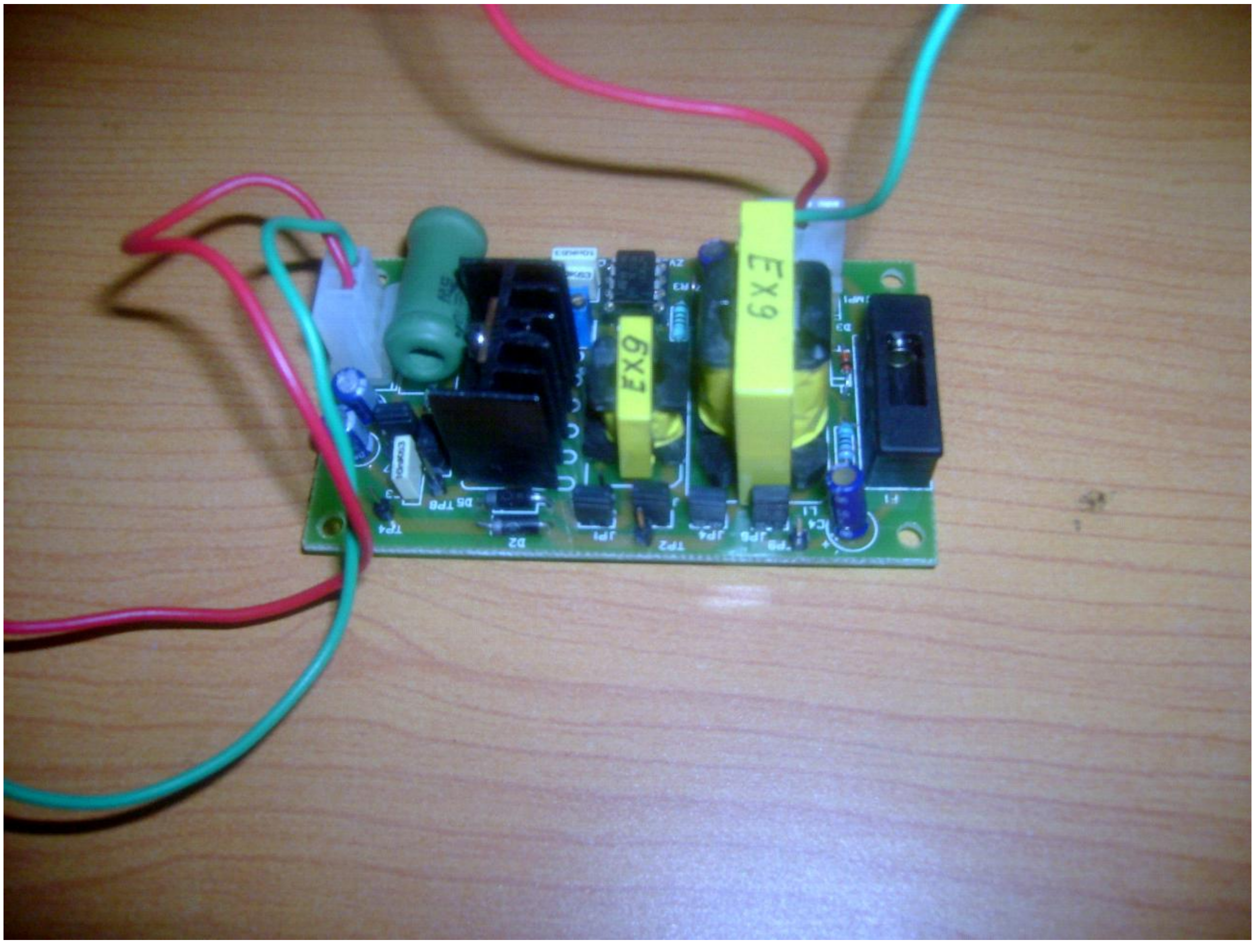
Fig 7.2 The Construction Project PCB

ȝTUCTION PROJECT 9 7.5W ZVS BOOST CONVERTER V2.0 03/02,

Bill Of Materials

SL. NO.	REFERENCE	DESCRIPTION	VALUE	QTY	RATE	AMOUNT
1	JMP1,JMP2	CONNECTOR	3 WAY PSC	2	3.00	6.00
2	R1	RESISTOR MFR	5K POT	1	7.00	7.00
3	R2	RESISTOR MFR	3.9K	1	0.35	0.35
4	R3	RESISTOR MFR	100E	1	0.35	0.35
5	R4	RESISTOR MFR	22E	1	0.35	0.35
6	R5	RESISTOR	33E,5W	1	5.00	5.00
7	R6	RESISTOR MFR	560E	1	0.35	0.35
8	R7	RESISTOR MFR	10K	1	0.35	0.35
9	C1	CAPACITOR	0.001uF/1nF/50V	1	1.10	1.10
10	C2,C3	CAPACITOR	0.01uF/10nF/63V	2	1.10	2.20
11	C4,C5,C6,C7	CAPACITOR	10uF/63V	4	0.50	2.00
12	D2,D5	DIODE	MUR110	2	8.00	16.00
13	D3	DIODE	10V,1/2W	1	1.00	1.00
14	F1	FUSE	2A	1	1.50	1.50
15	FUSE HOLDER	FUSE HOLDER	PBF4	1	2.80	2.80
16	Q1	TRANSISTER	2N2907	1	6.75	6.75
17	Q2	TRANSISTER	2N2222	1	6.75	6.75
18	Q3A	MOSFET	IRFZ44	1	16.00	16.00
19	U1	IC	555 TIMER	1	3.6	3.60
20	L1	INDUCTOR	E30/15/07,144T 24SWG,2mH	1	40.00	40.00
21	L2	INDUCTOR	E16/8/05,56.8uH 21T,24SWG	[illegible]	18.00	18.00
22	TP1,TP2,TP3,TP4,TP5,TP6 TP7,TP8,TP9,TP10,TP11	TEST POINT	1PIN BERG	[illegible]	0.10	1.10
23	JP1,JP2,JP3,JP4,JP5,JP6,JMP2	TEST POINT	2PIN BERG	7	0.20	1.40
24	HS2(25mm)	HEAT SINK	PI48/25MM	1	3.20	3.20
25	IC BASE	ROUND	8 PIN	1	2.20	2.20
26	TRANSISTER BASE		TO18	2	0.10	0.20
27	WIRING,CRIMPING			4	2.00	8.00
28	SHORTING LINKS			7	1.00	7.00
29	SCREWS		M3/8	2	0.11	0.22
30	PLANE WASHER		M3	1	0.10	0.10
31	FIBRE WASHER		M3	1	0.15	0.15
32	PACKING BOX			1	6.00	6.00
33	PCB-SS/SM/LP	IISC ZVS BOOST CONVERTER V2.0	7.5X6.0	1	22.00	22.00
						0.00

Table of Bill of materials



Constructed Circuit of the ZVS Boost Converter

Observations

The output dc voltages are observed as follows

Sl No.	INPUT DC (V)	OUTPUT DC (V)
1	10.5	16.2
2	11.2	18.4
3	12.0	22.2

The dc output for the rated voltage of 12V dc is observed on the CRO as shown in Fig 7.3. The Output of the 555 timer circuit having fixed R and C values is used for triggering the gate of the MOSFET. The Gate pulse waveform is also observed. The switching phenomenon can be seen between the Drain and the Source terminals of the MOSFET. The waveform for the same is shown in Fig 7.4. Various test point voltages are also recorded as shown.

OBSERVED WAVEFORMS

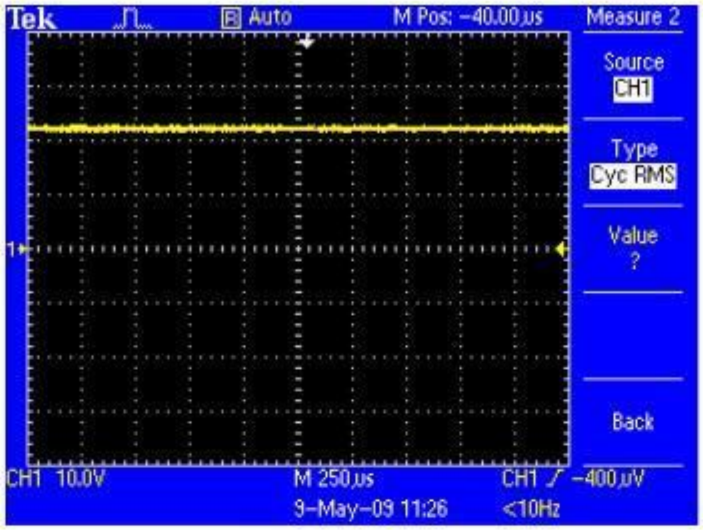


Fig 7.3 Output 22V DC for Input Voltage of 12V DC.

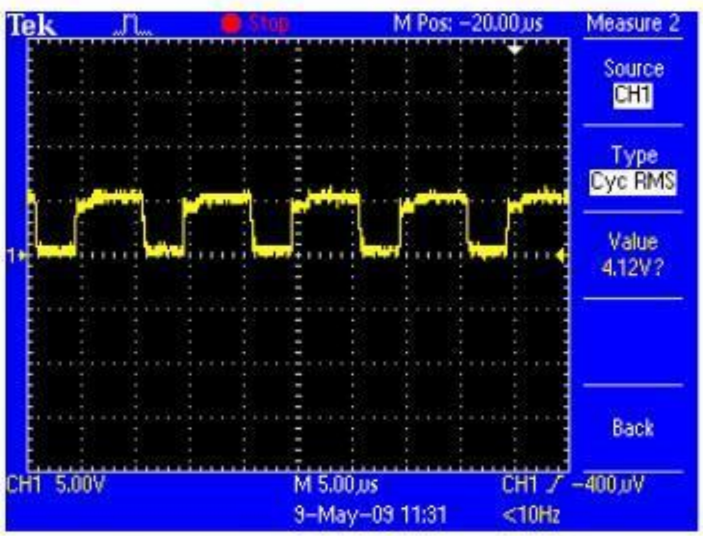


Fig 7.4 Source- Drain Switching Waveform

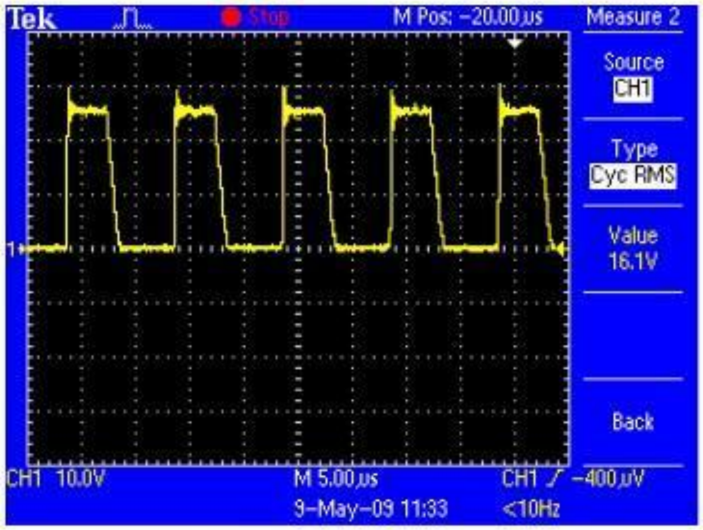


Fig 7.5 Gate Pulse Waveform

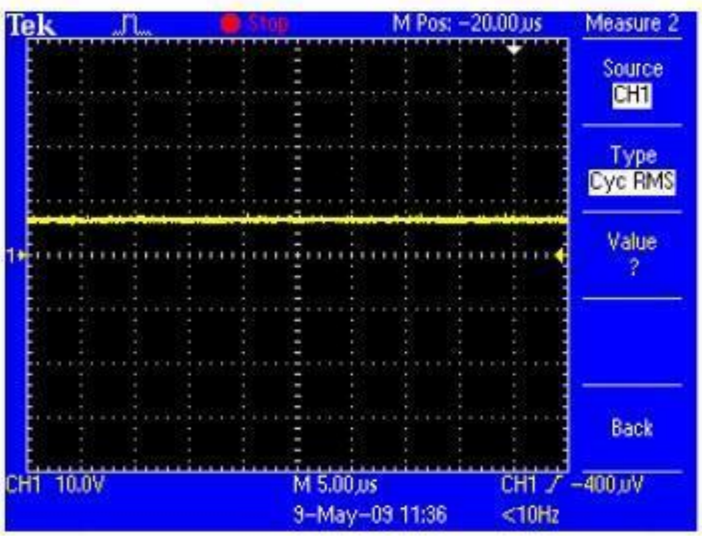


Fig 7.6 Test Point 1

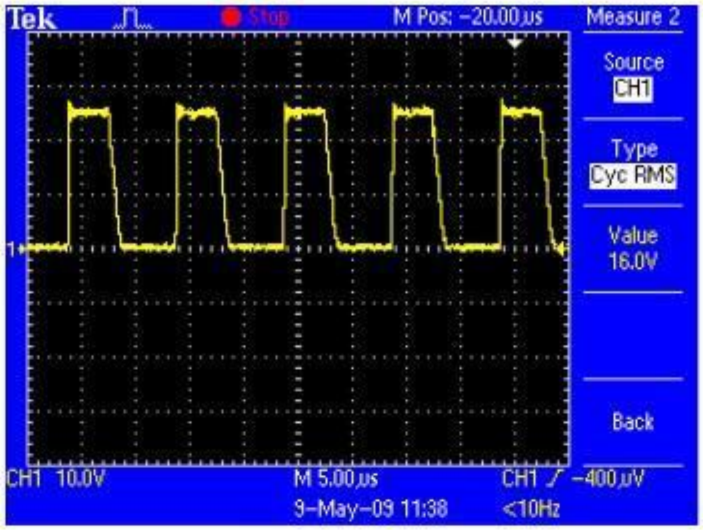


Fig 7.7 Test Point 2

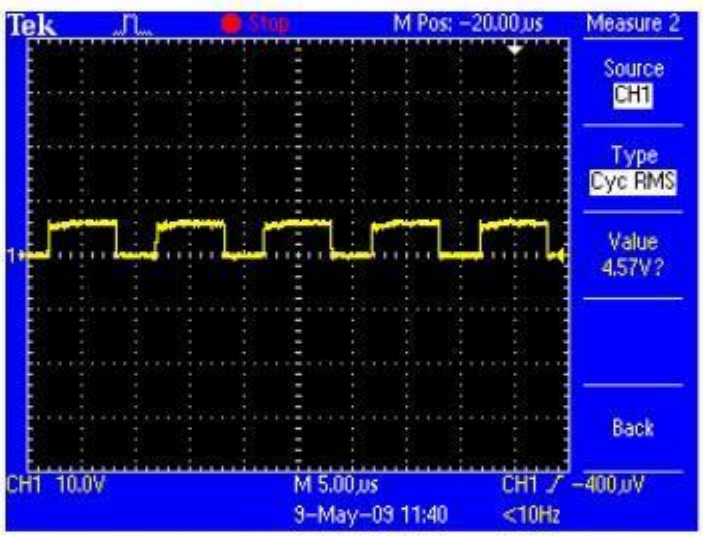


Fig 7.8 Test Point 3

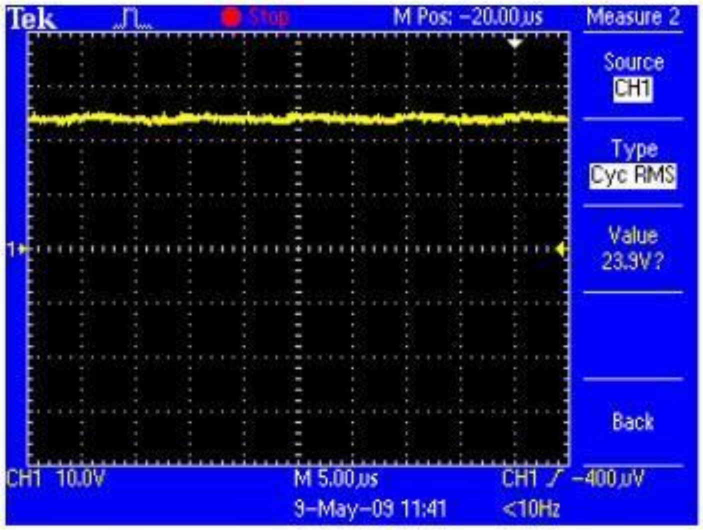


Fig 7.9 Test Point 4

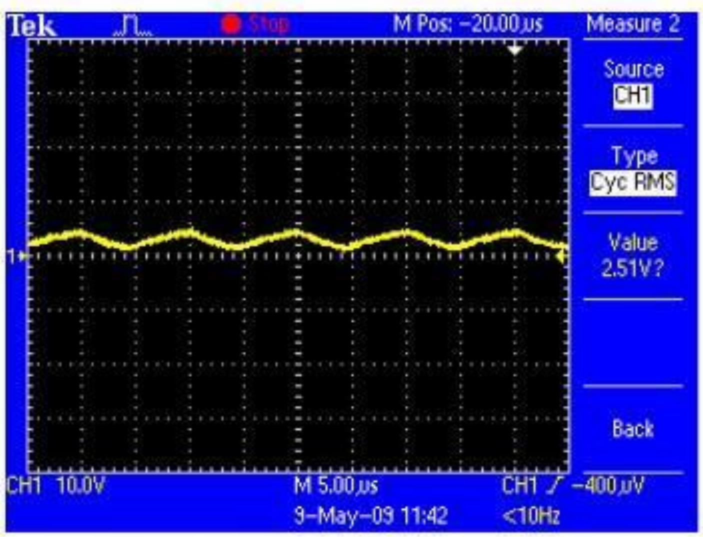


Fig 7.10 Test Point 5

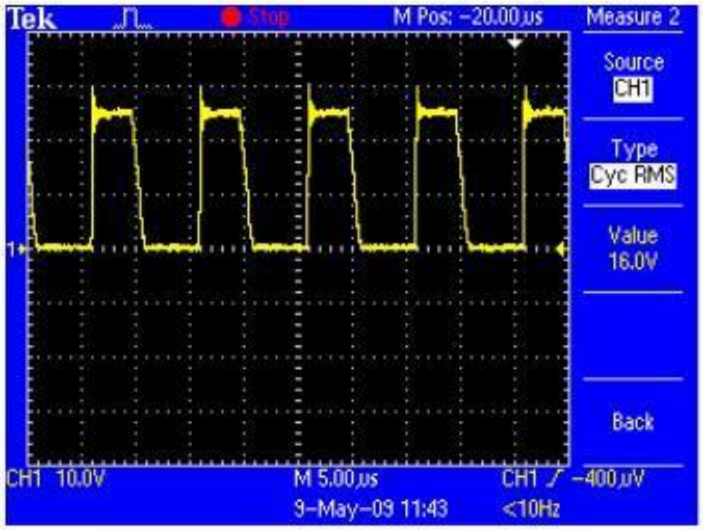


Fig 7.11 Test Point 7

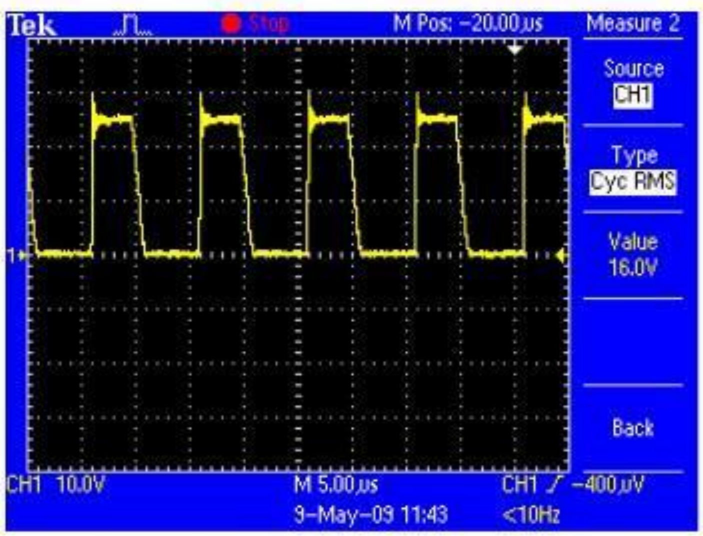


Fig 7.12 Test Point 9

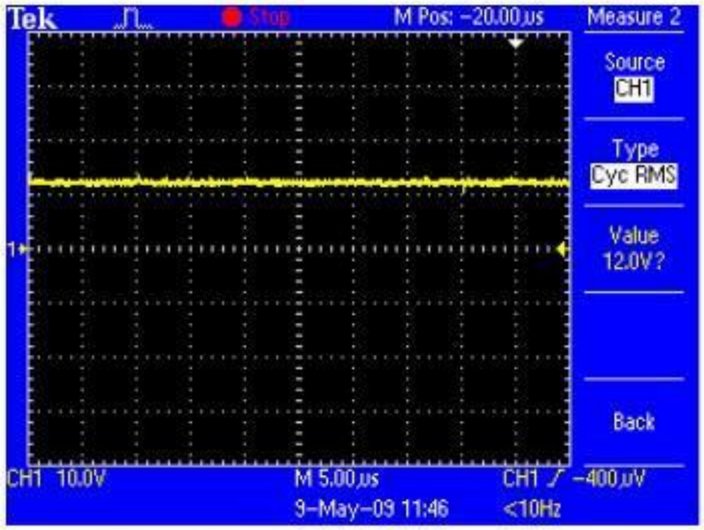


Fig 7.13 Test Point 10

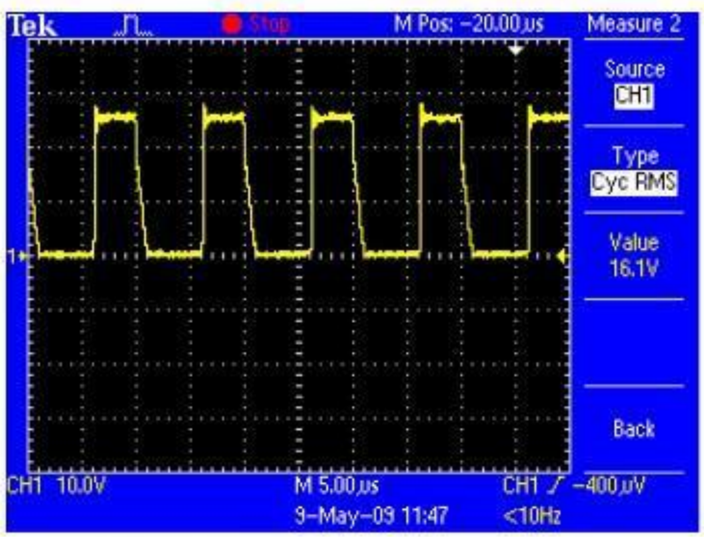


Fig 7.14 Test Point 11

CONCLUSION

ZVS Boost converter provides good zero voltage switching conditions for both the transistor and the diode. A ZVS Ciruit was realized and its waveforms were observed. Parasitic capacitance of the transistor and the diode parasitic inductances of connections are all parts of the resonant circuit. Switching of the transistor and the rectifying diode at zero voltage in the converter enables high operating frequency of the system while high energy efficiency is maintained. The range of the converter's operating frequency, in which ZVS switching is assured, is variable and dependent on the load resistance. ZVS boost converter generates dc voltage which can be applied in power supply systems where high energy efficiency is required.

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